RESEARCH DEPARTMENT

Variable-capacitance diodes as frequency multipliers

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RESEARCH DEPARTMENT

VARIABLE-CAPACITANCE DIODES AS FREQUENCY MULTIPLIERS

Technological Report No. G-094
(1964/60)

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VARIABLE-CAPACITANCE DIODES AS FREQUENCY MULTIPLIERS

SUMMARY

This report discusses the design of circuits using variable-capacitance diodes as high efficiency frequency multipliers with particular reference to their use at v.h.f. and u.h.f. Two multipliers are described, together with calculated performance and the practical results obtained.

1. INTRODUCTION

With the increasing use of the v.h.f. and u.h.f. bands for broadcasting and the more general use of semiconductors as active elements in equipment, the need has arisen for efficient means of generating power over the range 100 to 1000 Mc/s. At the present time the maximum powers obtainable with transistors vary from about 20 W at 100 Mc/s to about 30 mW at 1000 Mc/s.

Non-linear reactances may be used as efficient frequency multipliers, and capacitance (varactor) diodes when used as frequency doublers can give efficiencies of over 80% at high power or 60% at low power. The maximum powers obtainable with diode multipliers at the present time are in excess of 20 W at 300 Mc/s, 8 W at 1000 Mc/s and 1 W at 10 Gc/s. With these multipliers available, it is frequently advantageous to generate power efficiently at a comparatively low frequency with either a transistor or valve amplifier, and then follow this with high-efficiency diode multipliers to give the required output at the final frequency.

This method is of particular interest to the BBC as a means of generating several watts of r.f. power in Bands IV and V for transmitter site testing. Since the multipliers are compact and light it becomes practicable to fly a 5 to 10 W transmitter from a Mark VI kite balloon at heights of up to 1000 ft (305 m). With this low power and the use of sensitive receivers[1], it is possible to predict the service area of a 1000 kW station down to the 66 dB (relative to 1 μV/m) contour.

2. VARACTOR DIODES

When junction diodes are biased in the reverse or cut-off direction, they exhibit capacitance between the two electrodes. This capacitance varies with the applied voltage according to the law

$$C_v = K(V + V_0)^{\alpha}$$

where C_v is the capacitance at the applied voltage V

V is the applied voltage

K is a constant

V_0 is the contact potential (0·6 V for silicon)

α is a constant dependent on the method of construction of the diode.

For an abrupt-junction alloyed diode $\alpha \simeq -\frac{1}{2}$; for a gradual-junction diffused diode $\alpha \simeq -1/3$. The diodes used for frequency multiplication are normally of the alloyed type since they have a more non-linear capacitance characteristic than that of the diffused type, as indicated in Fig. 1.

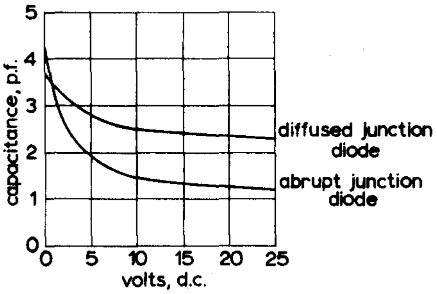

Fig. 1 - Typical diode characteristics based on manufacturer's data

The losses that can occur in varactor diodes are both series and shunt resistive losses. The shunt resistance of a diode is very high (a typical value at d.c. being 10,000 megohms), and can be entirely neglected. The series resistance of the diode is the bulk resistance of the silicon and is typically between 0·5 and 4 ohms. The resulting circuit magnifaction factor in tuned circuits of negligible inductive loss (i.e. the Q factor of the diode) may be above 1000 at 10 Mc/s, falling to 20 at 1000 Mc/s for a good diode.

3. SIMPLE FREQUENCY MULTIPLIER CIRCUITS

3.1. Bias Circuit

The diode in a multiplier must be supplied with a suitable bias potential, V_B; this may be either fixed bias or 'drive' bias produced by rectification of the applied signal, or it may be a combination of both. For maximum efficiency, the greatest variation in capacitance is required, and this occurs when the bias voltage has the minimum value which does not give conduction. This condition is not easy to maintain with fixed bias. If drive bias is obtained by permitting the diode to be driven just into conduction, considerable difficulty occurs in tuning the circuit and unstable operation may result. As the circuit approaches resonance the increased signal amplitude causes a rise in bias which changes the capacitance and hence the resonant frequency (see Fig. 2, curve a). If the drive is sufficient, the circuit can give discontinuities in its output levels (see Fig. 2, curve b). With some bias circuits the output may continuously oscillate between its two states.

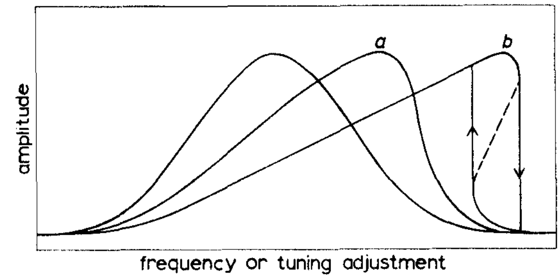

Fig. 2 - Variation of signal amplitude with frequency or tuning adjustment

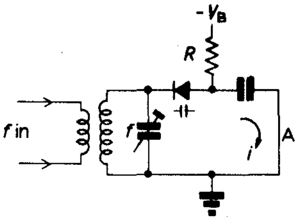

Fig. 3 - Basic series harmonic generator

In most practical multipliers it is usual to supply about two-thirds of the bias as a fixed bias and the remainder as drive bias. When a multiplier is driven hard, fixed bias is frequently used for setting up, but a drive bias circuit is used during operation. This has the advantage of preventing hum modulation from the bias supplies.

3.2. Series Type Harmonic Generator

In the circuit of Fig. 3 the tuned circuit is resonant at the fundamental frequency f. Ideally this tuned circuit would have an extremely high Q factor and very low impedance at all frequencies other than the fundamental. The total voltage across the varactor diode would then be

$$V_B + E \sin \omega t$$

where V_B is the bias voltage and $E \sin \omega t$ the voltage across the tuned circuit.

It can be shown[2] that if the diode has a perfect square law ($\alpha = -½$) and $V_o = 0$ then the current i through the diode is, subject to $E < V_B$,

$$i = \frac{2C.\omega E \cos \omega t}{\sqrt{(V_B + E \sin \omega t)}}$$

where C is the diode capacitance at 4 V bias.

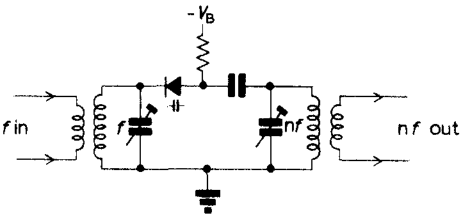

Fig. 4 - Practical series harmonic generator

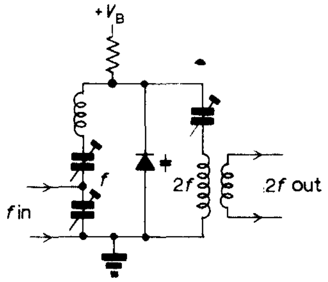

Fig. 5 - Practical shunt harmonic generator

For the condition of E approaching V_B, the current waveform is a highly distorted reproduction of the applied voltage, and it closely resembles a sawtooth. The formula for such a waveform may be expressed in the form

$$i = K(\sin \omega t + \frac{\sin 2\omega t}{2} + \frac{\sin 3\omega t}{3} + \ldots\ldots)$$

If a parallel tuned circuit is inserted at A, as in Fig. 4, power may be extracted at any harmonic frequency.

3.3. Shunt Type Harmonic Generator

The circuit shown in Fig. 5 is a dual of that of the series circuit and, in principle, similar output powers and efficiencies are obtainable.

3.4. Practical Considerations

In practical circuits there are a number of important differences which make one or other of the series or shunt generators preferable. In the series circuit the diode should have an impedance high compared with that of the tuned circuit components, whilst the shunt circuit requires a low impedance diode. In order to obtain the maximum available efficiency for a given diode the a.c. voltage across the diode should approach the breakdown voltage. This means a high impedance diode is preferred for very low powers, whilst high powers require a low impedance diode.

Since all harmonic currents flow in the series circuit, these currents will give additional losses. The shunt circuit, however, produces only harmonic voltages and harmonic currents will not flow unless a suitable path is provided. This gives rise to less loss in the shunt circuit than in the series circuit. With the shunt circuit it is usually necessary to provide parallel-tuned traps, as shown in Fig. 6, to reduce direct feed-through from input to output and vice versa.

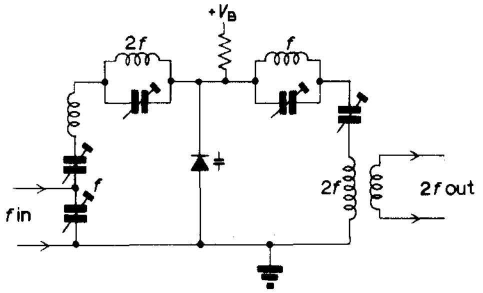

Fig. 6 - Shunt Circuit with tuned traps

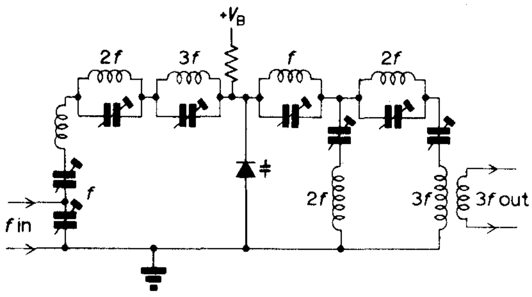

Fig. 7 - Shunt trebler circuit with second harmonic idler

Corresponding traps in the series circuit give a slight improvement in performance. If harmonic outputs of higher order than the second are required, the series circuit will provide them with a comparatively small loss in efficiency. The shunt circuit, however, requires series tuned 'idler' circuits at the intermediate harmonics, together with the appropriate rejection filters as shown in Fig. 7. These idler circuits, being unloaded, will have a high Q factor and will be difficult to tune, especially when more than one idler is necessary. For this reason the simple shunt circuit is not normally used for any multiplication other than doubling.

By a suitable choice of input and output circuits the necessity of using traps may be reduced. For example, in Fig. 5, the input is applied across a capacitor which reduces harmonic currents in the input whilst the associated inductor maintains a high impedance to the harmonics in this arm of the circuit. Conversely, the output circuit uses a small capacitor of high impedance at the fundamental frequency in series with the output whilst a small inductor across the output further reduces direct break-through of the fundamental. In some cases this inductor may be partially tuned by a parallel capacitor to give even greater rejection.

The series circuit is normally used for low-power doublers and all multiplications greater than four times in a single stage.

For high-order multiplications it is preferable to use a series of doublers or treblers rather than one single multiplier. When cascading these harmonic generators, critically coupled double-tuned circuits should be used between stages to reduce any possibility of interstage coupling at unwanted harmonic frequencies, since such coupling may lead to much reduced efficiency and may produce spurious outputs which are difficult to remove by filtering.

4. BROADBAND AND HIGH-POWER CIRCUITS

In the harmonic generators described in Section 3, tuned traps are required for good efficiency. These lightly loaded tuned circuits will have a relatively small bandwidth. By using a balanced circuit these traps can be eliminated and much larger bandwidths obtained[3]. Since the balanced circuits use two diodes, they have a higher power handling capacity and the provision of heat sinks is simplified because one side of the diodes may be earthed.

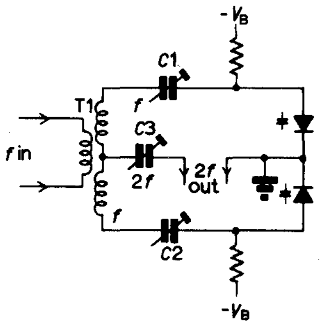

Fig. 8 - A balanced doubler

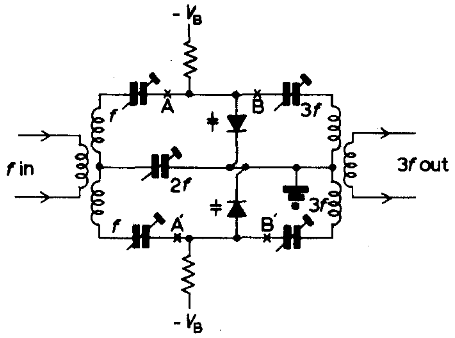

Fig. 9 - A balanced tripler

The balanced doubler of Fig. 8 has a balanced input and gives an unbalanced output. The trimmer capacitors C1 and C2 match the capacitances of the diodes which tune the transformer T1 at the fundamental frequency. The trimmer C3 series tunes the large leakage inductance of the transformer T1 at the second harmonic. The output is taken from this series circuit. Since the circuit is symmetrical, no fundamental can appear in the output, nor can any second harmonic appear in the input. The even harmonics produced by the diodes add in the output circuit.

If circuits tuned to the third harmonic are added across the diodes, as shown in Fig. 9, and the second harmonic load short-circuited, a third harmonic output may be obtained. Parallel tuned traps similar to those of Fig. 7 will be required at AA' BB' for good efficiency and a minimum level of spurious frequencies, but will give a reduction in bandwidth.

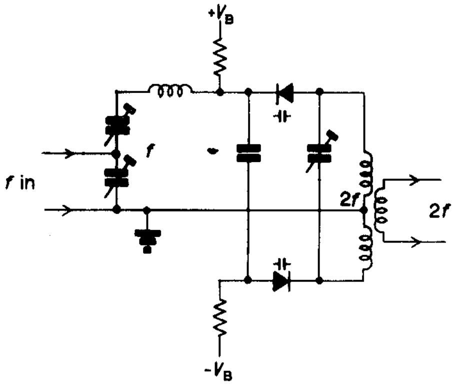

Fig. 10 - A hybrid doubler

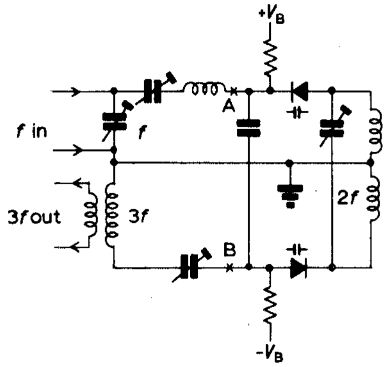

Fig. 11 - A hybrid tripler

The hybrid doubler circuit of Fig. 10 uses an unbalanced input and balanced output. The input is series tuned and the output parallel tuned. It has the advantage compared with the circuit of Fig. 8 that the tuned circuits are independent and consequently are easy to adjust. A series tuned output may be used but is much more difficult to tune. The chief disadvantage of this circuit is that it is difficult to fit efficient heat sinks to the diodes because they are at a high r.f. potential.

By adding a circuit tuned to the third harmonic, as in Fig. 11, a trebler is obtained. Again, traps may be required at A and B for optimum performance. The unloaded idler circuit at the second harmonic limits the bandwidth of the circuit. If this idler circuit is short-circuited, as in Fig. 12, the circulating second harmonic currents give an extra loss of about 10% but the bandwidth is increased considerably. If several stagger-tuned traps are used at A and B, bandwidths in excess of 10% may be obtained.

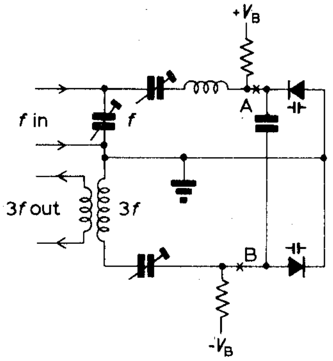

Fig. 12 - Wide bandwidth tripler

5. BANDWIDTH

In circuits without traps or idler circuits, bandwidths of up to 30% may be obtained on doublers or treblers. Tuned traps will normally limit the bandwidth to less than 5% although this may be increased by the use of multiple stagger-tuned traps. Idler circuits will usually limit the bandwidth to less than 1%.

6. MODULATION

6.1. Frequency Modulation

Within the bandwidth of the multiplier, frequency modulation is unaffected except that its deviation is increased by the multiplication factor.

6.2. Amplitude Modulation

A diode harmonic generator can be amplitude modulated in an 'efficiency' mode by varying the bias; the modulation law is, however, very non-linear.

If used as a doubler with an amplitude modulated input, it gives considerable second harmonic distortion of the modulation and, even when the circuit bias is adjusted for best linearity, the linearity is only fair (5% approximately) above 20% of the maximum power output and the overall efficiency is poor.

Where it is necessary to obtain amplitude modulation, a varactor diode may be used as a parametric up-converter[4,5] in which an amplitude modulated signal at a comparatively low frequency is amplified and raised in frequency.

The dynamic range of this type of converter may exceed 140 dB, which enables it to be used as an exceptionally linear up-converter and amplifier.

6.3. Pulse Modulation

If a pulsed input is applied to a multiplier, the distortion tends to sharpen the leading and trailing edges of the pulse. The reduced dissipation in the diodes can give some increase in efficiency and power handling capacity compared with a c.w. input.

7. OVERDRIVE

If any of the circuits of Figs. 4 to 12 are overdriven, the diodes overheat, their leakage currents increase, and the drive and output circuits become detuned. The circulating current then falls, and the diode drive reduces, making the circuit go into a thermal oscillation with a time constant dependent on the diode and heat sink. A diode without a heat sink will usually oscillate at about 4 c/s. It is very difficult to damage a diode permanently by any reasonable overdrive.

8. NOISE

Varactor diode multiplier chains produce very little noise provided that the bias supply is adequately filtered. In microwave receivers a klystron local oscillator can be replaced by such a chain with a reduction of local oscillator noise[6].

9. TEMPERATURE

Temperature changes in normal environments have a negligible effect on diode multipliers unless they are working at an exceptionally low level. However, the temperature coefficient of the tuned circuits themselves must be adequately compensated or wideband circuits used, otherwise a fall in efficiency may occur through detuning.

10. POWER AND EFFICIENCY

The energy stored in a linear capacitor C when charged to a voltage V_B is $\frac{1}{2} C V_B^2$. The volt-amperes that can be transferred to the load via a varactor diode is proportional to the product of this energy and the frequency[7].

Thus, if the peak drive voltage equals the bias voltage the maximum volt-amperes or reactive power that can be transferred will be given by

$$VA = \tfrac{1}{2} C V_B^2 . K f_{in} \qquad (1)$$

where C is the capacitance at the bias voltage V_B

f_{in} is the input frequency

K is a constant with appropriate dimensions. It is dependent on circuit design.

The usual value of K is between 0·5 and 1, but values up to 10 may be obtained in efficient multipliers.

The cut-off frequency f_c of a diode is defined as the frequency for which the series resistance R_s of the diode is equal to the reactance X_c of the diode. The addition of this resistance in series with the diode will give a loss term of the form $I^2 R_s$ at all frequencies at which currents flow. This includes input and output frequencies, together with any other harmonic currents which may be present.

The efficiency η is given by

$$\eta = \eta_0 \, \eta_1 \left(1 - \alpha_n \frac{f_{out}}{f_c}\right) \qquad (2)$$

where η_0 is the efficiency in the absence of all resistive losses.

η_1 is the proportionate efficiency reduction factor due to tuned circuit losses.

f_{out} is the output frequency

f_c is the cut-off frequency where $R_s = X_c$

α_n is the dimensionless constant for harmonic n and has been found empirically[7] to have a value of approximately $5n$ for small values of n.

The difference between η_0 and unity is a measure of the generation of unwanted harmonics. Simple doubler circuits permit $\eta_0 > 0{\cdot}9$.

At frequencies up to 1000 Mc/s, provided that good diodes at sufficiently high power are used, losses of n dB can usually be obtained for the n^{th} harmonic without undue difficulty.

The above formulae are approximate for low levels of drive voltage. In practice it is found that the drive may be increased above the value required to drive the diode into conduction, to give an appreciable increase of output power at a slightly lower efficiency. Too large an increase in drive, however, results in a rapid increase of losses.

11. PRACTICAL CIRCUITS

In this section two high-power frequency multipliers which have been constructed are described. Consideration is given to theoretical details and practical results.

Multiplier chains using varactor diodes have been used to provide the local oscillator in u.h.f. field strength recording and measurement receivers[1] and also in the construction of a v.h.f. standard field oscillator which is used for the calibration of field strength measurement aerials and receivers by the 'standard field' method[8].

The first multiplier to be described is a single-diode frequency sextupler, multiplying from 91·67 Mc/s to 550 Mc/s. As expected, this multiplier had a high level of spurious outputs, but it showed that a high output could be obtained from a single stage.

The second multiplier to be described is a frequency doubler from 91·65 Mc/s to 183·34 Mc/s for use as a Band III pilot transmitter for site testing.

11.1. Frequency Sextupler (91·67 Mc/s to 550 Mc/s)

The main purpose of this multiplier was to ascertain the maximum power obtainable from a single diode with a multiplication factor of six and to determine what difficulties were likely to occur in the construction. Idler circuits at $2f$ and $3f$ were used[9,10]. The circuit is given in Fig. 13 and the experimental model is illustrated in Fig. 14. The diode type PC117 (Pacific Semiconductors) when used with a suitable heat sink, will dissipate 4 W, although considerable care in design and in mounting the diode is required. The diode has a maximum working voltage of 100 V, a capacitance of 15 pF at 50 V and a series resistance of approximately 1/3 ohm.

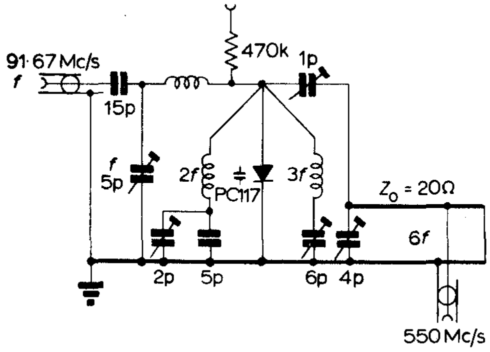

Fig. 13 - Circuit of sextupler

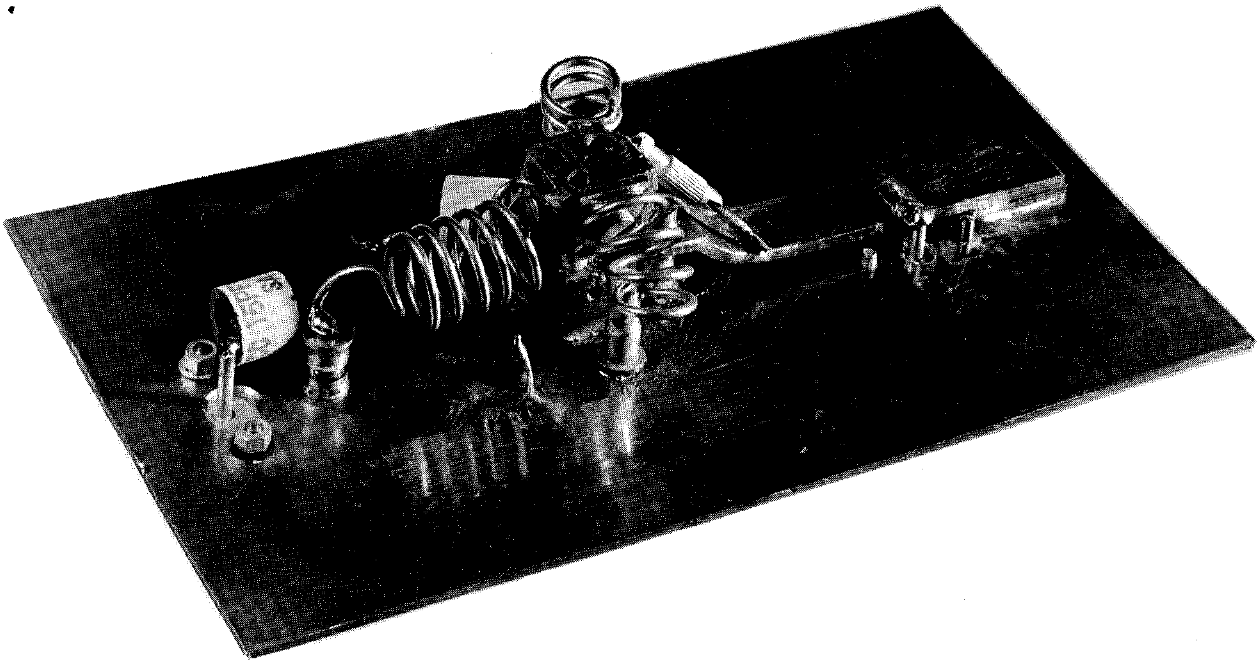

Fig. 14 - Photograph of sextupler

From Section 10 equation (1) the maximum reactive power ($W_D^{\bullet}$) transferred may be deduced as 6·9 volt-amperes.

11.1.1. Efficiency

The loaded Q factors will all be less than 1/30 of the unloaded Q factors and it is reasonable to assume, in the absence of series resistance, an efficiency η_1 of approximately 0·9. Since the unwanted harmonic generation should be small, $\eta_0 = 1$. Also $\alpha_6 = 30$ and, for the particular diode in use, ($R_s = 1/3$ ohms and $C = 15$ pF) $f_c = 30$ Gc/s. These values give an overall efficiency from equation (2) of $\eta = 0{\cdot}40$, i.e. a transfer loss of 4 dB.

Assuming that the efficiency of the input tuned circuit is 0·97, then the required input power for maximum output is 6·9/0·97 = 7·12 W and the output power will be 0·4 x 7·12 = 2·85 W.

The dissipation in the diode, from the appropriate term of equation (2) is

$$\alpha_n \frac{f_{out}}{f_c} \cdot W_D$$
$$= 0{\cdot}55 \times 6{\cdot}9 \text{ W}$$
$$= \quad 3{\cdot}8 \text{ W}$$

Since the optimum fundamental voltage at the diode is the full diode rating, i.e. 100 V peak-to-peak or 35 V r.m.s., the current will be 6·9/35 = 0·197 amps, and the load resistance of the diode at the fundamental frequency will be 178 ohms.

At 6*f* the voltage will be one sixth (5·83 V) and the current will be 6·9 x 0·4/5·83 = 0·472 amps. Hence the output resistance of the diode will be 12·2 ohms.

As suitable components were not to hand when this unit was built, an input circuit of very low loaded Q was used. This permitted about 1/3 W at 2*f* (183·34 Mc/s) to be lost in the drive unit and gave a poor match to the circuit. This loss reduces η_o to 0·9, giving an overall loss of 4·4 dB and an output power of 2·5 W.

Idler circuit components were chosen to give loaded Q factors of 5 and 7 respectively, thus giving low losses and reasonable values of capacitance and inductance for the tuned circuits. The output circuit matches the diode (output resistance) to the output line. The circuit used for this consists of a parallel tuned $\lambda/8$ line (Z_o = 20 ohms) with a small series capacitor to the diode. The output is taken from a tap on the line giving a loaded Q factor of 70. This should reduce all unwanted harmonics of the input by at least 20 dB. Fixed bias is used for setting up.

11.1.2. Measurement on Sextupler Circuit

It was found on test that the input to the multiplier could be increased above the calculated maximum of 7·1 W. The maximum input was 10·5 W, giving an output of 3·8 W at 550 Mc/s, corresponding to a loss of 4·4 dB. If an attempt was made to increase the drive above this value, the input match deteriorated rapidly, giving a slightly reduced output.

The main spurious outputs were at

458·33 Mc/s	(x 5)	approximately 0·2 W	(−13 dB)
641·67 Mc/s	(x 7)	approximately 0·2 W	(−13 dB)
1008·37 Mc/s	(x 11)	approximately 0·3 W	(−11 dB)

All other multiples up to x 10 gave outputs in the range 1 to 50 mW (−36 dB to −19 dB). The high outputs at x 5 and x 7 were apparently due to strong coupling between the idler and output circuits. The high output at x 11 was due to the mechanical arrangement necessitated by the type of trimmer used to couple the output circuit to the diode. This gave a spurious resonance at this frequency which could easily have been avoided by a slightly different design.

The experiment showed that a single diode sextupler could be made to give a reasonable output power but for practical purposes considerable filtering would be required to reduce unwanted harmonics to an acceptable level.

11.2. Frequency Doubler (91·7 Mc/s to 183·4 Mc/s)

In order to carry out Band III site tests it became necessary to produce a balloon-borne transmitter for use at a fixed frequency within this band. In order to reduce development effort, an existing 91·7 Mc/s balloon-borne transmitter[11] was used,

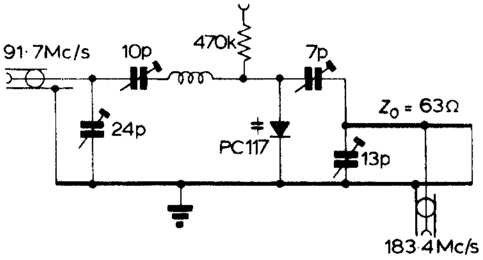

Fig. 15 - Circuit of doubler

together with a simple varactor diode doubler (Figs. 15 and 16). With a measured bias V_B of 65 V the calculated maximum output for this doubler is 10·2 W for an input of 12·0 W, a loss of 0·7 dB. In practice the doubler gave an output of 16 W for an input of 25 W, corresponding to a loss of 1·8 dB. The diode dissipation at this power level is excessive for c.w. operation but as the transmitter is square-wave modulated the diode dissipation is satisfactory, the mean output power being 8 W. A simple stub (λ/4 at 91·7 Mc/s) across the output feeder reduces all spurious outputs to less than 10 μW.

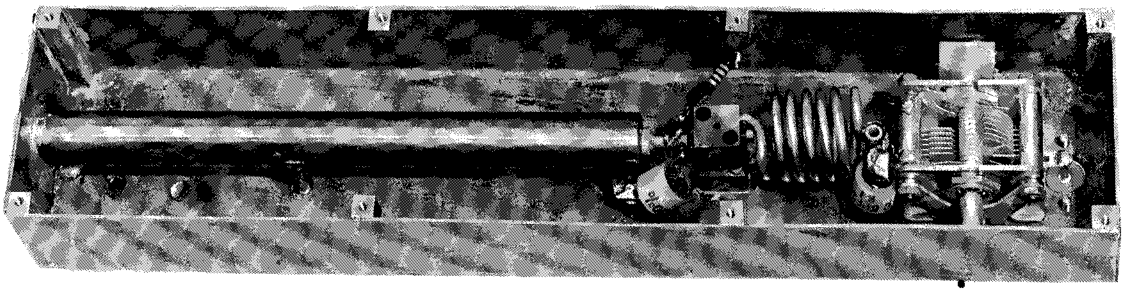

Fig. 16 - Photograph of Band III doubler

12. CONCLUSIONS

Low-loss varactor diode multipliers can be designed to give powers of at least 10 W at frequencies in the range 200 to 1000 Mc/s. The powers obtainable in practice are usually slightly greater than the calculated values but at the expense of a slightly greater loss in the circuit. Thus an increase of input power above the calculated optimum is required.

Unwanted harmonics produced by the diode can be held to low levels but only if care is taken to avoid stray couplings and resonances.

It is relatively easy to design low-level multipliers suitable for use as local oscillators in v.h.f. and u.h.f. receivers.

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